VI. Communications Elements Research

ADVANCED ENGINEERING

A. Low Noise Receivers: Microwave Maser Development, *E. R. Wiebe*

1. Introduction

An instrument for automatically recording the reserve heat capacity of traveling-wave maser (TWM)/closed cycle refrigerators (CCR) has been built and tested at JPL. Using this device, the overall health of CCR systems in the field can be easily monitored between tracking periods. The measurement of reserve heat capacity is performed with higher accuracy and greater speed than possible with the manual methods generally in use.

2. Instrument Description

The instrument consists of a simple servo system which applies sufficient dc power to the liquid helium stage to maintain its temperature at about 0.1°K above normal. This power level is then displayed as reserve heat capacity on either a meter or strip chart recorder.

Figure 1 shows a schematic diagram of the unit. It can be divided into three parts:

(1) The power supply.

(2) The control unit.

(3) The indicator unit.

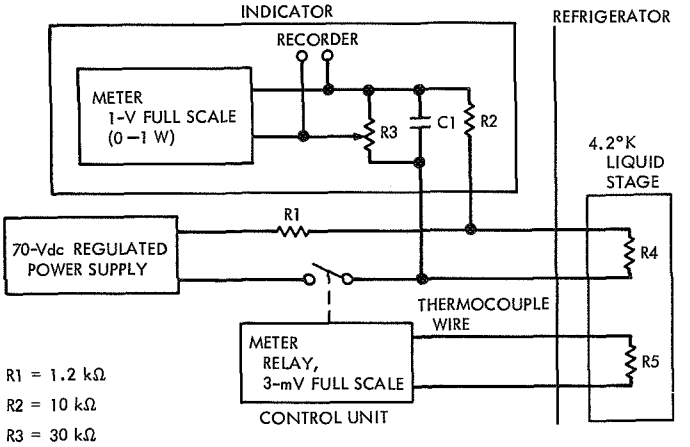

Fig. 1. Instrument for recording reserve heat capacity of TWM/CCR system

The power supply is adjusted to provide 1 W of power to heater resistor R4 on the liquid helium stage. R1 minimizes inaccuracies due to different values of heater resistance from one refrigerator to another. The control unit is a meter relay connected in series with a thermocouple and temperature-sensitive resistor R5. The meter relay is adjusted to switch the power supply on-and-off at a rate which maintains the liquid helium stage temperature at about 0.1°K above normal. The indicator circuit is a meter with a 0- –1-W linear scale connected to a voltage divider R3 and a resistance/capacitance (R/C) integrator R2 and C1. R3 is adjusted for a full-scale meter reading when a continuous 1 W is applied to the CCR heater R4. The R/C integrator provides a voltage to the meter which is proportional to the percentage of time the heater is switched on, therefore, giving a direct reading of dc watts reserve heat capacity.

3. Instrument Operation

Figure 2 shows a typical recording of reserve heat capacity from a CCR used to cool a TWM. At $T = 0$ h, after the refrigerator has cooled to operating temperature, the instrument described in *Subsection 2* was energized. Supply pressure (SP) and Joule-Thompson (J-T) pressure were both 300 psi. At these pressure conditions, the refrigerator had a J-T flow of 1.18 standard ft³/min (scfm) and a reserve heat capacity of about 880 mW.

To demonstrate how the instrument works, the above pressures were reduced to 250 and 225 psi, respectively. At $T = 2$ h, the capacity is seen to drop to around 600 mW. At about $T = 4$ h, the maser pump klystron was turned on and capacity was dropped to about 540 mW, indicating about 60 mW of pump energy directly absorbed by the maser.

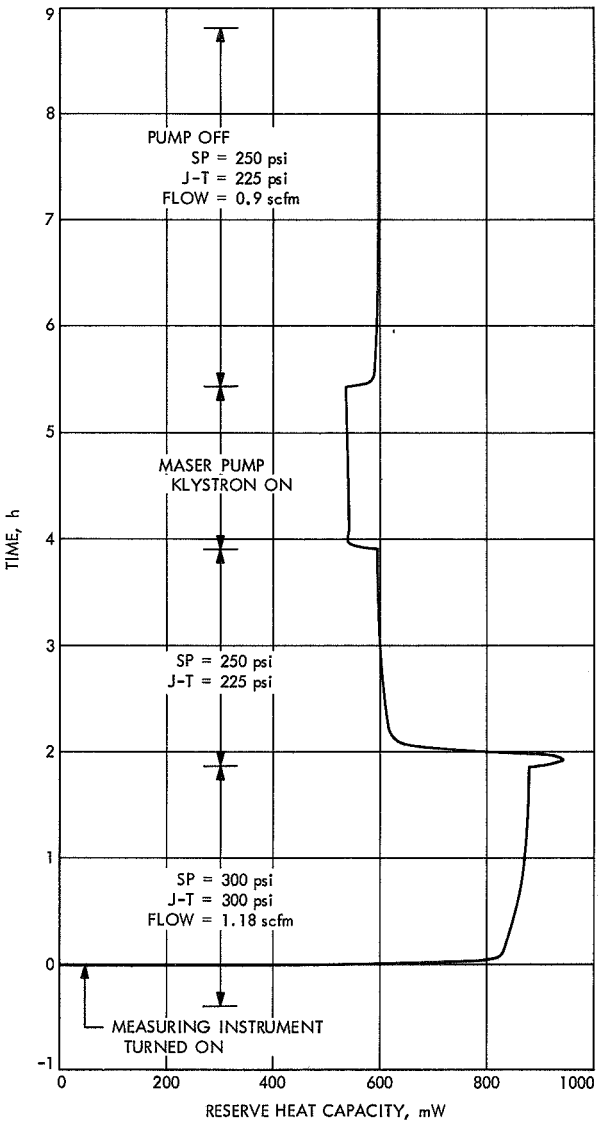

Fig. 2. Recording of reserve heat capacity of TWM/CCR system

B. Improved Calibration Techniques: Realizability Conditions on Reflection Coefficients of Unsymmetrical, Passive, Reciprocal 2-Port Networks, *T. Y. Otoshi*

1. Introduction

It is generally known that if a 2-port reciprocal microwave network is lossless, then

$$|S_{11}| = |S_{22}|$$

where $|S_{11}|$ and $|S_{22}|$ are the magnitudes of the voltage reflection coefficients as seen looking into the input and output ports, respectively, when the opposite port is terminated in a non-reflecting load.

In practice, however, microwave networks are not lossless. For a practical network that is lossy and reciprocal (but not symmetrical), it is often desired that an estimate of $|S_{22}|$ be obtained from knowledge of $|S_{11}|$. Heretofore, a simple formula for obtaining this estimate has not been widely known or available in published form. Recently, the necessary formulas have been derived by Dr. Glenn F. Engen of the National Bureau of Standards, and it is the purpose of this article to discuss these formulas and point out their potential usefulness to microwave engineers.

2. Formulas for Limits of $|S_{22}|$

During a technical discussion with Dr. Glenn F. Engen, it was discovered that bounds on $|S_{22}|$ in terms of $|S_{11}|$ could be derived from the following realizability relationship, which appeared in a previous paper by Engen (Ref. 1).

$$1 - \frac{|S_{12}|^2}{1 - |S_{11}|^2} \geqq \left| S_{22} + \frac{S_{11}^* S_{12}^2}{1 - |S_{11}|^2} \right| \tag{1}$$

where S_{11}, S_{12}, and S_{22} are scattering coefficients and (*) denotes complex conjugate.

If reciprocity is assumed, i.e., $S_{21} = S_{12}$, then from Eq. (1), the following limits of $|S_{22}|$ can be derived in terms of $|S_{11}|$.

$$|S_{22}|_{\max} = \frac{1}{L}|S_{11}| + \left(1 - \frac{1}{L}\right) \tag{2}$$

If $|S_{11}| > L - 1$, then

$$|S_{22}|_{\min} = \frac{1}{L}|S_{11}| - \left(1 - \frac{1}{L}\right) \tag{3a}$$

If $|S_{11}| \leqq L - 1$, then by additional physical considerations

$$|S_{22}|_{\min} = 0 \tag{3b}$$

where

$$L = \frac{1 - |S_{11}|^2}{|S_{21}|^2} \tag{4}$$

The symbol L is a power ratio $\geqq 1$ and is often referred to as the dissipative loss factor when the direction of energy flow is from port 1 to port 2. Letting $L_{\text{dB}} = 10 \log_{10} L$, one can use the following low-loss case approximations:

$$\left.\begin{aligned} \frac{1}{L} &\approx 1 - 0.23\, L_{\text{dB}} \\ L - 1 &\approx 0.23\, L_{\text{dB}} \\ 1 - \frac{1}{L} &\approx 0.23\, L_{\text{dB}} \end{aligned}\right\} \quad \text{for } L_{\text{dB}} \leqq 0.1 \text{ dB} \tag{5}$$

In cases where the attenuation (or the matched system insertion loss) of the 2-port is known instead of L_{dB}, the dissipative loss factor can be calculated from

$$L = (1 - |S_{11}|^2)\, 10^{A_{\text{dB}}/10} \tag{6}$$

where A_{dB} is the network attenuation in decibels.

3. Applications

Three practical situations where the $|S_{22}|$ limit formulas can be useful are described below.

(1) In the microwave calibrations of a reciprocal 2-port network where the waveguide types on each port are different, it may not be economical to assemble reflectometer systems for each waveguide type. If the loss of the 2-port device is known to be small, then from Eqs. (2) and (3), it might be found that the difference between $|S_{22}|$ and $|S_{11}|$ will be negligibly small. If such is the case, it is only required that reflection coefficient measurement be made at one of the ports.

(2) It is sometimes the case where reflection coefficients have been measured at both ports of a reciprocal 2-port network, but due to a possible measurement error, the data obtained for one of the ports appears to be questionable. Since the measured values must be bounded by the physically realizable limits, the limit formulas can be used to check the validity of the measurement data.

(3) In conjunction with the design of a new microwave system, a theoretical study is often made of the effect of such sources of errors as mismatches which can cause significant deviations from predicted system performance. In such studies, it becomes necessary to study hypothetical cases and synthesize worst-case system parameter values. Equations (2) and (3) can be used to insure that physically realizable values of voltage reflection coefficients are used in the hypothetical test cases.

4. Sample Cases

The following are a few sample cases which demonstrate the spreading that occurs between maximum and minimum values of $|S_{22}|$ as the lossiness of the 2-port is increased. Comparisons are made with actual measured values of $|S_{22}|$.

(*1*) *Lossless case.* In the lossless case, the 2-port has no dissipative losses so that

$$L = 1$$

and the limits from Eqs. (2) and (3) become

$$|S_{22}|_{max} = |S_{22}|_{min} = |S_{11}|$$

(2) *Sample low-loss case.* Given that

$$L_{dB} = 0.0016\,dB$$

$$|S_{11}| = 0.00798$$

then from use of Eqs. (2), (3), and (5),

$$|S_{22}|_{max} = 0.00835$$

$$|S_{22}|_{min} = 0.00761$$

The input data of this case was obtained from measurements made on a WC504/WR 430 copper broadband transition. The actual measured value of $|S_{22}|$ was 0.00847 $\pm$0.00023 pe_T (Ref. 2).

(3) *Sample medium-loss case.* Given that

$$A_{dB} = 0.172\,dB$$

$$|S_{11}| = 0.0004$$

then from Eqs. (2), (3), and (6),

$$|S_{22}|_{max} = 0.0392$$

$$|S_{22}|_{min} = 0$$

The input data of this case was obtained from measurements made on a cryogenic load transmission line of an *X*-band Cassegrain cone system. The actual measured value of $|S_{22}|$ was 0.021 (SPS 37-43, Vol. III, p. 59).

(*4*) *Sample high-loss case.* Given that

$$A_{dB} = 1.17\,dB$$

$$|S_{11}| = 0.0075$$

then from Eqs. (2), (3), and (6),

$$|S_{22}|_{max} = 0.242$$

$$|S_{22}|_{min} = 0$$

The input data of this case was obtained from measurements made on a compact *H*-band rotary vane attenuator whose vane was set at 20 deg. The actual measured value of $|S_{22}|$ at this setting was 0.015 (SPS 37-62, Vol. II, pp. 81–87).

5. Conclusions

It has been shown that when the loss of a 2-port network is known and the reflection coefficient is measured at only one of the ports, the limits of the magnitude of reflection coefficient at the other port can be easily calculated. It is important to remember that the limit formulas presented in this article are restricted to passive 2-port networks that are reciprocal (non-ferrous).

References

1. Engen, G. F., "An Evaluation of the Back-to-Back Method of Measuring Adaptor Efficiency," *IEEE Trans. Instr. Meas.*, Vol. IM-19, No. 1, pp. 18–22, Eq. (10), Feb. 1970.

2. Levy, G. S., et al., "The Ultra-Cone: An Ultra-Low Noise Space Communication Ground Radio Frequency System," *IEEE Trans. Micro. Theory Tech.*, Vol. MTT-16, No. 9, pp. 596–602, Sep. 1968.

C. A Noise-Adding Radiometer for Use in the DSN,

P. D. Batelaan, R. M. Goldstein, and C. T. Stelzried

1. Introduction

The communication requirements for deep space missions necessitate the use of high-gain antennas and sensitive receivers on the ground. Because of their unusually good sensitivity, these receiving systems are also well suited to radio astronomy applications. However, most gain compensation techniques normally employed for radio astronomy, such as the Dicke scheme, would degrade the communication capability of a DSN receiver. A radiometer which is compatible with the DSN communication requirement is described here.

2. Discussion

The radiometer scheme chosen is based on a technique due to Ohm and Snell (Ref. 1) except that a digital computer is used to handle the output data rather than the analog device of Ohm and Snell. A block diagram of the noise-adding radiometer (NAR) is shown in Fig. 1. Figure 2 depicts the portable NAR rack and its contents.

A solid-state noise diode (noise temperature ~400,000 K) is driven by a constant current source. The input current to the diode is square-wave-shorted by computer control at a low-frequency keying rate (11 Hz, typically). The resulting square-wave noise is coupled (−40 dB, typically) into the receiver input. This is a noise temperature T_N increase of ~ 40 K added into the receiver input at an 11-Hz square-wave rate. This is depicted in Fig. 3. The noise temperature of the receiver is multiplied by the overall gain G and it is the instability of this gain term which is eliminated to achieve increased radiometer sensitivity.

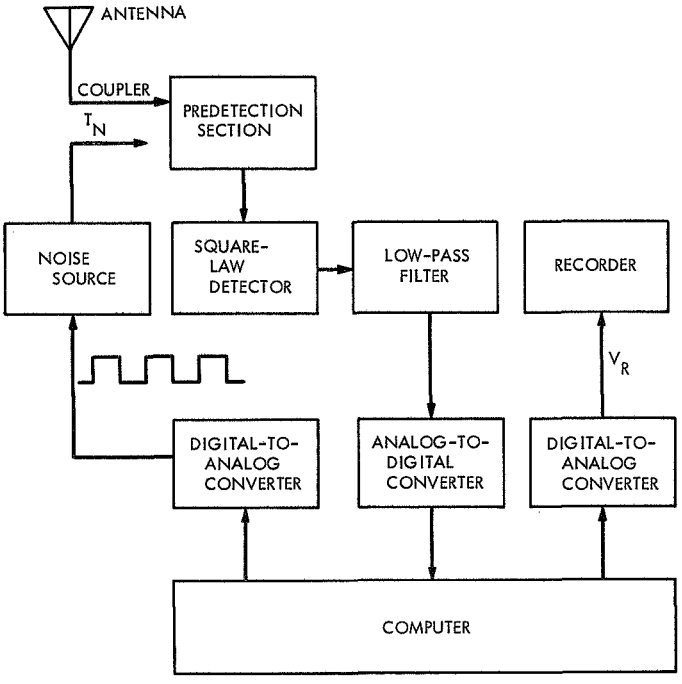

Fig. 1. Noise-adding radiometer

Salient to the operation of this NAR is a good, wideband square-law detector. The computer periodically (i.e., typically 7 kHz) samples the output of the square-law detector and forms a voltage ratio (Fig. 3)

$$V_R = \frac{v_1}{v_2 - v_1} = \frac{G(T_{op} + T_S)}{GT_N} \tag{1}$$

where

v_1 = averaged voltage at the square-law detector output during noise diode off cycle

v_2 = averaged voltage at the square-law detector output during noise diode on cycle

T_{op} = system operating noise temperature

T_S = radio source temperature

Rearranging, we have

$$\frac{T_S}{T_N} = V_R - \frac{T_{op}}{T_N} \tag{2}$$

which is gain independent. An initial zeroing operation is made with the antenna off source. A term $(T_{op}/T_N)_0$, the value of V_R when $T_S/T_N = 0$, is stored in the computer and subtracted from subsequent outputs (i.e., typically every 0.6 s). The computer output is then given by

$$V_R = \frac{T_S}{T_N} + \left[\frac{T_{op}}{T_N} - \left(\frac{T_{op}}{T_N}\right)_0\right] \tag{3}$$

This output is linearly related to the source temperature. The second term is zero immediately after the zeroing operations and drift is negligible during operation. Instability in the noise diode is reflected as corresponding instability in the radiometer.

Noise calibration of the system is accomplished by injecting a separate noise-source signal into the receiver input. The value of the calibration source is 1–4 K (typical) referenced to the antenna.

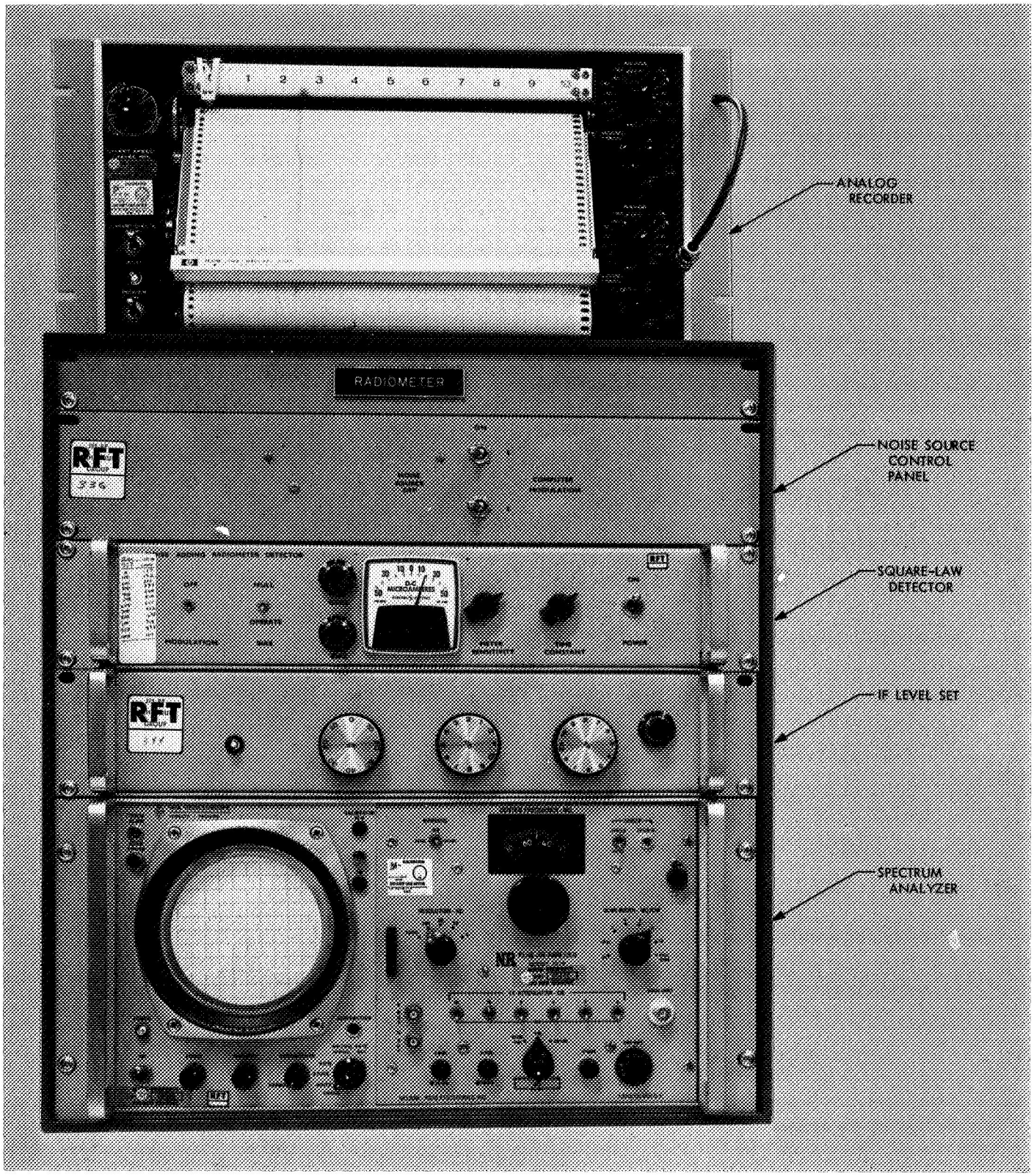

Fig. 2. Noise-adding radiometer rack

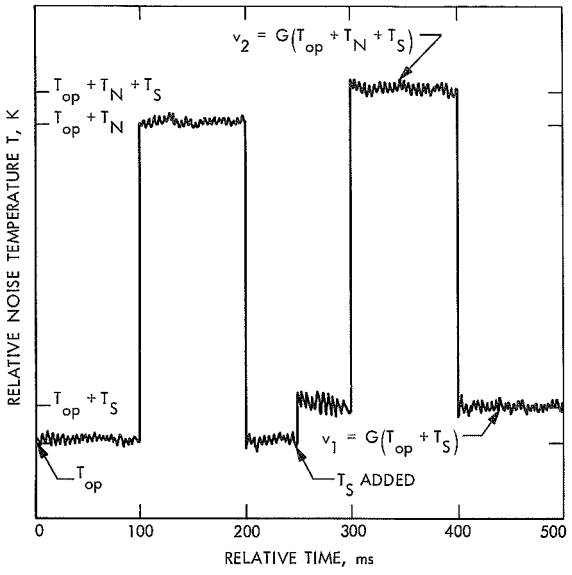

Fig. 3. Representation of the signal levels from the square-law detector

A comparison of the NAR with a total power radiometer (TPR) will serve to clarify their relative characteristics. The TPR sensitivity is given by (Ref. 2)

$$(\Delta T_{\min})_{\mathrm{TPR}} = T_{op}\left[\left(\frac{1}{\Delta\nu_{HF}\tau_{LF}}\right) + \left(\frac{\Delta G}{G}\right)^2\right]^{1/2} \tag{4}$$

where

$\Delta\nu_{HF}$ = detector bandwidth, Hz

τ_{LF} = equivalent integration time of the output integrator

G = system gain

and

ΔG = change (instability) in G

Without gain instability (the ideal radiometer),

$$\Delta T_{\min} = T_{op}\left(\frac{1}{\Delta\nu_{HF}\tau_{LF}}\right)^{1/2} \tag{5}$$

The NAR sensitivity is (*Subsection 4*)

$$(\Delta T_{\min})_{\mathrm{NAR}} = T_{op}\,2\left(\frac{1}{\Delta\nu_{HF}\tau_{LF}}\right)^{1/2}\left(1 + \frac{T_{op}}{T_N}\right) \tag{6}$$

Two observations can be made: (1) if the gain is constant for the TPR and injected noise is high for the NAR, then

$$(\Delta T_{\min})_{\mathrm{NAR}} = 2\,(\Delta T_{\min})_{\mathrm{TPR}} \tag{7}$$

and (2) the NAR exhibits (theoretically) no sensitivity to radiometer gain instability.

The NAR is not yet fully operational but some routine star tracks have been executed in the course of system development. Results have been excellent. Figure 4 shows a simultaneous TPR and NAR recording of source 3C10.

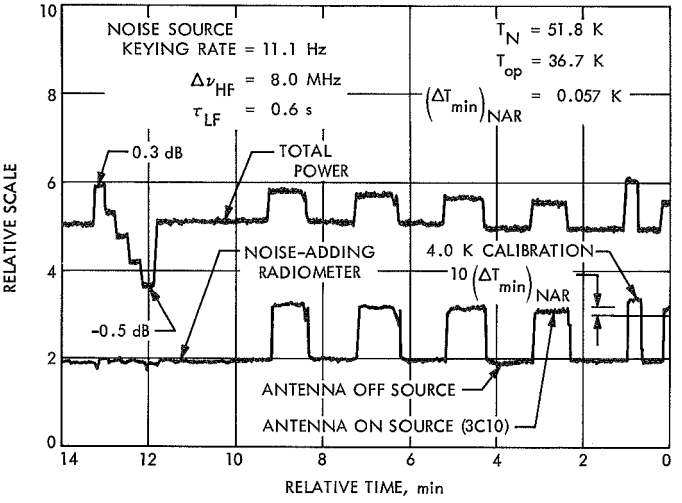

Fig. 4. Noise-adding radiometer recording (DSS 11, Feb. 11, 1970)

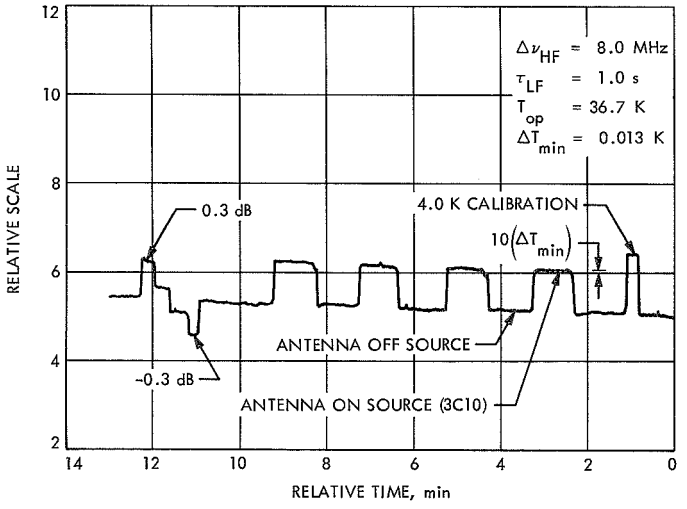

Fig. 5. Total power radiometer recording (DSS 11, Feb. 11, 1970)

While the antenna was pointing on the background sky, system gain was deliberately changed. Notice that the NAR was relatively unaffected while the TPR output was radically changed. The total power radiometer used for recording Fig. 4 has degraded sensitivity since it was operating simultaneously with the NAR and its associated noise injecting source. Figure 5 shows a recording of the normal (NAR not in operation) operation of the TPR. Again, notice the gain sensitivity due to deliberate gain changes and also due to drift.

3. Conclusion

Examination of Fig. 4 shows that the NAR was operating with a minimum sensitivity of 0.066 K, which is about 1.2 times the theoretical minimum of 0.057 K determined from Eq. (6). Similarly, Fig. 5 shows that the TPR was operating with a minimum sensitivity of 0.019 K, which is about 1.5 times its theoretical (ideal) minimum of 0.013 K determined from Eq. (5).

The conclusion to be drawn from Figs. 4 and 5 is that the total power radiometer has lower short-term jitter and is most useful for short-term measurements (seconds) and that the noise-adding radiometer has less drift and is most useful for long-term measurements (minutes).

The noise-adding technique of achieving a gain insensitive radiometer has been demonstrated to be a powerful technique which is ideally suited to application at a DSN station. Care must be taken in fabrication of a true square-law detector and the usual problem of gain linearity must be controlled throughout the system. The keyed noise-source stability is also critical.

4. Calculation of $\Delta T_{\min}$

From Fig. 3, the average values of v_1 and v_2 are seen to be

$$v_1 = G(T_{op} + T_S) \tag{8}$$

$$v_2 = G(T_{op} + T_N + T_S) \tag{9}$$

For errors Δv_1 and Δv_2, we have from Eq. (1), since $|\Delta v_1| << v_1$ and $|\Delta v_2| << v_2$,

$$\Delta V_R = \left(\frac{\partial V_R}{\partial v_1}\right)\Delta v_1 + \left(\frac{\partial V_R}{\partial v_2}\right)\Delta v_2 \tag{10}$$

Since v_1 and v_2 are independent (bar denotes average),

$$\sigma^2_{V_R} = (\Delta V_R)^2 = \left(\frac{\partial V_R}{\partial v_1}\right)^2 \overline{(\Delta v_1)^2} + \left(\frac{\partial V_R}{\partial v_2}\right)^2 \overline{(\Delta v_2)^2}$$

$$= \frac{v_2^2\,\overline{(\Delta v_1)^2} + v_1^2\,\overline{(\Delta v_2)^2}}{(v_2 - v_1)^4} \tag{11}$$

Let P be the switching cycle period. Then we have a total power radiometer during each half switching cycle and from Ref. 2, p. 245,

$$\overline{(\Delta v_i)^2} = \frac{v_i^2}{(\Delta \nu_{HF})\frac{P}{2}}, \qquad i = 1, 2 \tag{12}$$

Using Eqs. (11) and (12), we obtain

$$\sigma_{V_R} = \frac{2v_1v_2}{(v_2 - v_1)^2\,(\Delta \nu_{HF} P)^{1/2}} \tag{13}$$

The minimum detectable temperature $\Delta T_{\min}$ may be defined as that T_S which causes V_R to change by σ_{V_R}. From Eq. (2), the ΔV_R due to $T_S = \Delta T_{\min}$ is

$$\Delta V_R = \frac{\Delta T_{\min}}{T_N} \tag{14}$$

Upon equating Eqs. (13) and (14) and using Eqs. (8) and (9) with T_S suppressed because it is small compared to T_{op},

$$\Delta T_{\min} = \frac{2\,T_{op}\left(1 + \frac{T_{op}}{T_N}\right)}{(\Delta \nu_{HF} P)^{1/2}} \tag{15}$$

This was derived for one switching period. Clearly, if τ_{LF} is any integer number of switching periods, we may replace P by τ_{LF} because we simply have an average over a longer time. Thus,

$$(\Delta T_{\min})_{\mathrm{NAR}} = \frac{2\,T_{op}\left(1 + \frac{T_{op}}{T_N}\right)}{(\Delta \nu_{HF} \tau_{LF})^{1/2}} \tag{16}$$

References

1. Ohm, E. A., and Snell, W. W., "A Radiometer for a Space Communications Receiver," *Bell System Technical Journal*, pp. 2047–2080, Sep. 1963.

2. Kraus, J. D., *Radio Astronomy*. McGraw-Hill Book Co., Inc., New York, 1966.